CELP CODING FOR LAND MOBILE RADIO APPLICATIONS

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ABSTRACT

Described is the performance of the Proposed Federal Standard 1024 8000 bps system for Digital Land Mobile Radio in a 12.5 or 6.25 kHz channel.

The system contains the Proposed Federal Standard 1016 4800 bps Code Excited Linear Predictive (CELP) voice coder developed by the U.S. DOD and ATT Bell Laboratories. The system is error protected with 2400 bps and has 800 bps for overhead signaling. The system in the future will replace the 12 kHz Continuous Variable Slope Deltamod (CVSD) voice coder which is used in the present Federal Standard 1025.

The error protection technique employed is Golay coding with hard or soft decoding. We describe the performance of this error protection over simulated fading radio channels.

1. INTRODUCTION

The U.S. government is currently developing Federal Standard 1024 which will specify the baseband transmission format, voice coder, error protection, and modulation to be used in future digital land mobile radio applications with 12.5 or 6.25 kHz channel spacing. Present land mobile radio is based on 25 kHz channels, but increasing demands on the fixed spectrum availability along with overseas evolution towards narrower channels have caused the need for more efficient spectrum use. The bit rate of the FS1024 system is 8000 bps, consisting of 4800 bps voice coder, 2400 bps Golay [1] code error protection on the most sensitive voice bits, and 800 bps overhead for system control and synchronization. The modulation selected for this study was pi/4 shifted QDPSK, the same as that adopted for future digital cellular radio. The voice processor, error correction and modulation are integrated together for optimum performance. The system performance is evaluated using radio channel simulators with emphasis given to the performance of the Golay code over burst error conditions normally seen on land mobile radio channels.

2. CELP ALGORITHM DESCRIPTION

CELP is a frame-oriented voice coding technique that breaks a sampled speech signal into blocks of samples which are processed as one unit [2]. CELP is based on analysis-by-synthesis search procedures, two-stage perceptually weighted vector quantization (VQ), and linear prediction. A 10th order linear prediction filter is used to model the speech signal's short-term spectrum and is commonly referred to as a spectrum predictor. Long-term signal periodicity is modeled by an adaptive code book VQ (also called pitch VQ because it often follows the speaker's pitch in voiced speech). The residual from the spectrum prediction and pitch VQ is vector quantized using a fixed stochastic code book. The optimal scaled excitation vectors from the adaptive and stochastic code books are selected by minimizing a time-varying, perceptually weighted distortion measure. The perceptual weighting function improves subjective speech quality by exploiting masking properties of human hearing.

CELP uses an 8 kHz sample rate and a 30 ms frame size with four 7.5 ms subframes. CELP analysis consists of three basic functions : 1) short term spectrum prediction, 2) long delay adaptive code book "pitch" search, and 3) innovation stochastic code book search. CELP synthesis consists of the corresponding three synthesis functions performed in reverse order with the optional addition of a postfilter to enhance the output speech. The transmitted parameters (and bits per frame) for this model are the stochastic code book index (9x4) and gain (5x4), the adaptive code book index (8+6+8+6) and gain (5x4), and 10 line spectral parameters or LSP (34), sync (1), error correction (4), and future expansion (1), yielding a frame size of 144 bits for the coder.

3. CELP ERROR PROTECTION

Parameter coding and continuity are the basis for our error protection strategy internal to the 4800 bps CELP stream. We developed an integrated adaptive error protection system combining forward error correction (FEC), smoothers and parameter coding. Adaptive smoothers are employed, based on estimates of the channel error rate derived from the time averaging of the FEC code syndrome detection. This allows the smoothers to be disabled in error-free conditions. Efforts were concentrated on preventing perceptually disturbing synthesis errors such as loud speech blasts and squeaks. Pitch delay is the CELP parameter most perceptually sensitive to errors after smoothing all appropriate parameters. For this reason, 4 bits of FEC from a Hamming (15,11) single error detecting and correcting code are used to protect 10 pitch delay and pitch gain bits. Pitch gain has many nonsmooth regions where smoothing is ineffective, therefore we protect its most significant bit. The 11th bit protected by the Hamming code is the future expansion bit. With this internal protection strategy, CELP operates reasonably well in bit error rates on the order of 1%.

Land mobile radio channels are likely to subject the CELP bit stream to average error rates of up to 10%, caused by continuous rates of up to several percent due to operation of the LMR in geographic margins where Gaussian noise predominates, and burst rates of 50% due to fades below the noise floor. Due to these poor operating environments it was seen necessary to add 2400 bps error protection to the basic CELP stream. The FEC selected was the half-rate (24,12) Golay encoder and is used to protect the 72 most sensitive bits in the 144 bit CELP frame. This code can correct 3 and detect 4 errors per code word. Studies are ongoing as to which of the CELP bits will be protected. Initial indications are that these should include the Hamming parity bits, the formerly unprotected pitch delays and gains, and the LSPs. In order to alleviate the effects of fades wider than CELP frames, output speech frames will be repeated when a certain number of the 6 Golay code words are detected to have 4 or more

errors. Frame repeats will likely be allowed up to 2 consecutive frames, with output speech squelching being employed to span very long fades.

4. SYSTEM SIMULATION

The simulated system is shown in Fig 1. Speech data was transmitted for evaluation of speech quality over the simulated radio channels, whereas pseudorandom PN data was transmitted for error rate and distribution analysis. Separate PN sequences were used for the Golay code word (CW) information bits and the uncoded (X) bits. The overhead bits consisted of a 24 bit random sequence used for frame sync purposes at the receiver. In a realized FS1024 system they would represent the 800 bps overhead described earlier.

The modem used was a pi/4 shift QDPSK, operating at 4000 bauds /sec with sampling rate of 12000 samples/sec and carrier at 3000 Hz (channel simulators were baseband). Candidates for FS1024 modulation include 4-ary FSK, tamed FM, QDPSK and pi/4 shift QDPSK. At the time of paper preparation, the final selection had not been made, but will be predicated on spectral efficiency for adjacent channel interference minimization and power efficiency.

Two channel simulators were employed. SIM1 was a simplified burst fade simulator that had two modes - fading during which the signal was divided by 200 and subjected to -24 dB S/N, and non-fading during which the S/N could be any level. The fade width and rate statistics were controlled. SIM2 was a sophisticated VHFUHF software radio channel simulator based on a 3-tap chain differentiator model. This simulator could be used to simulate Rayleigh fading channels with any desired level of rms multipath spread or doppler spread, or Rician fading channels with up to 10 independent paths with controllable amplitude, delay, and doppler shift.

Interleaving is utilized to battle the effects of fades. Fig 3 shows the interleaving table employed. The enclosed block indicates the CELP information bits. The input matrix to the interleaver is at top and the output matrix is at bottom, with the output bit stream read left-to-right and top-to-bottom. Code words are spread maximally in the output sequence.

The integrated nature of the error control strategy is highlighted in Fig 1 by the indication of the bit confidence metrics being passed from the demodulator through the deinterleaver to the Golay decoder. In order to achieve the performance gain possible by the use of soft decoding, this additional link is necessary. These confidence values are real values as are the sampled signals going in and out of the channel simulators. All other links in Fig 1 comprise bits. The confidence metrics are computed as shown in Fig 2. The indicated dot shows the unquantized phase difference between the present baud and previous baud, and the magnitude of the present baud relative to the low-pass filtered receive baud magnitude represented by the circle. The transmitted bauds (10, 00, 01, 11) correspond to phase changes (45, 135, -135, -45 degrees). The X and Y axes are the lsb and msb decision boundaries. The lsb confidence is simply the phase distance from the nearest lsb boundary and likewise for the msb. These two confidence values are scaled by the amplitude so as to flag signal fades. Soft Golay decoding employs these confidence metrics by finding the lowest 4 confidence bits in a code word. The 16 bit patterns possible (2**4) are in turn used to invert the designated bits in the code word which is then Golay decoded. If the result is error-free then it is accepted. If no candidate is error-free, the detected errors are added to the test error pattern and the cumulative confidence is computed. Finally, the bit pattern candidate with the lowest cumulative confidence is selected. Up to 7 errors can be corrected.

5. PERFORMANCE & RESULTS

Initial studies were directed at which type of QDPSK detection was to be employed in the demodulator - Coherent or Noncoherent. As can be seen in Fig 4, coherent outperforms noncoherent on S/N channels on SIM1. As would be expected, in both cases the soft outperforms the hard which in turn outperforms the Golay turned off. These results may imply that coherent detection is preferred, but results in Fig 10 employing SIM2 show that the phase disruption caused by doppler spread clearly favored the noncoherent. In the coherent case no attempt was made to track the carrier phase, with the receiver instead being given the known transmitted carrier phase. The received absolute phase constellation (after pi/4 deshifting) wandered all over the 360 degrees instead of centering on the desired 4 points. If a coherent detector were to be implemented that could track the SIM2 phase irregularities then the test data would improve. As it is, all test data in the remainder of the paper will be from the noncoherent detector, and thus conservative.

Using SIM1, a channel was implemented that simulated what INMARSAT considers to be typical of land mobile radio [3]. The characteristics of this channel are 90% availability with 10% fades of durations (10, 20, 40, 100, 200 msec) and corresponding relative probabilities (.8, .1, .05, .04, .01). Fig 5 shows the results with the uncoded error rate at .057, hard at .062, and soft at .038. The results are clearly dominated by the 100 msec fade which spans more than 3 CELP frames. This is a probable squelching situation. In order to get better statistics, SIM1 was used to parameterize both fade width and fade rate. Fig 6 shows the results with fade width varied, with 10 msec representing by far the most predominant fade width in the INMARSAT channel recommendation. As can be seen, the soft decoder clearly helps this case. However, at wider fades the coder collapses since it can only "chip away at the edges". Fig 7 shows the time waveforms of the burst width signals. The fade rate parameterization results are shown in Fig 8 and Fig 9, where the basic INMARSAT fade width of 10 msec is employed throughout.

Using SIM2, flat fading doppler spread channels were simulated. Multipath spread of up to 5 usec was also simulated with results about the same as with 0 usec flat fading. Fig 11 and 12 show the results with doppler spreads of 30, 60 and 100 Hz (each with noise floors of 12, 15 and 18 dB S/N where the signal level is the overall rms). The phase constellations shown in Fig 12 are the unquantized phase differences in 12 dB S/N. The results are encouraging, with the 60 Hz, 18 dB test yielding BERs of uncoded=.023, hard=.0035, and soft=.0002. The error distributions are shown in Fig 13. Note also the improvement of the Golay decoder as the doppler spread increases, due to the fade durations decreasing. Multipath Rician channels were also simulated using SIM2. The results are shown in Fig 14 and 15.

6. CONCLUSIONS

From the test results, it is evident that the addition of a half rate soft Golay FEC to the CELP bit stream of the FS1024 will yield highly improved performance. How the postulated radio channels will impact the uncoded CELP bits is yet to be determined. These bits are less essential to acceptable CELP performance, and should be able to tolerate a higher error rate. Future work will entail coherent detection QDPSK, processing of a large quantity of speech through the simulated radio channels, and simulation of a more varied collection of radio channels.

The apparent limitation of the Golay technique is the obvious one of fade width toleration. Certainly as fade widths become too wide, schemes such as frame repeat and squelching become necessary.

7. ACKNOWLEDGEMENTS

We are grateful for our in-house support from : John Lee, Rich Dean, Ron Cohn, Tina Kohler, and Dave Kemp.

8. REFERENCES

[1] A. Levesque and A. Michelson, *Error-Control Techniques for Digital Communication*, Wiley, 1985, pp. 157-170.

[2] J. Campbell, V. Welch and T. Tremain, *The New 4800 bps Voice Coding Standard*, Mil Speech Tech Conference, November 1989.

[3] S. Wong, *INMARSAT Codec Evaluation Process ICEP Working Document ICEP/WK/1*, August 1989.

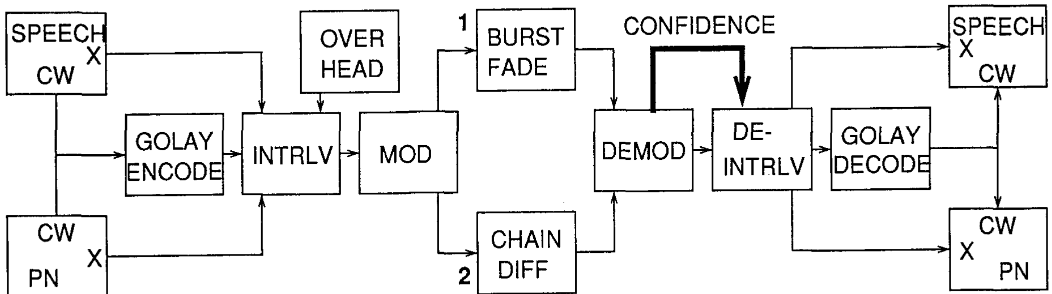

Fig 1. BLOCK DIAGRAM OF SIMULATED SYSTEM

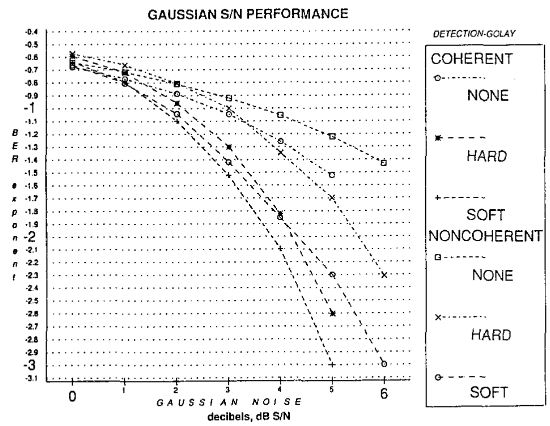

Fig 4. GAUSSIAN NOISE PERFORMANCE

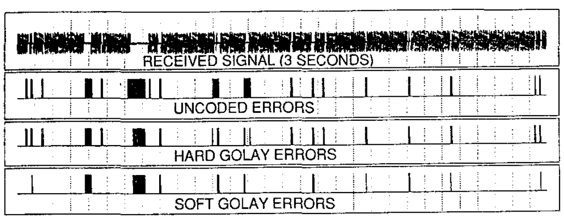

Fig 5. INMARSAT CHANNEL PERFORMANCE

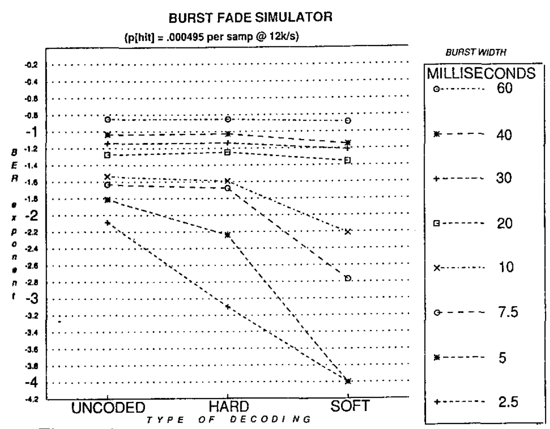

Fig 6. VARYING FADE WIDTH PERFORMANCE

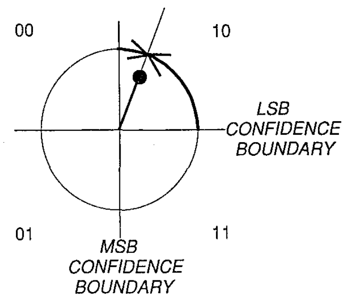

Fig 2. CONFIDENCE METRIC COMPUTATION

F	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
CW1	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
CW2	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
CW3	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
CW4	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
CW5	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
CW6	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
X1	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
X2	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
X3	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24

F	CW1	CW2	X1	CW3	CW4	X2	CW5	CW6	X3
1	1	1	1	1	1	1	1	1	1
2	13	13	2	13	13	2	13	13	2
3	2	2	3	2	2	3	2	2	3
4	14	14	4	14	14	4	14	14	4
5	3	3	5	3	3	5	3	3	5
6	15	15	6	15	15	6	15	15	6
7	4	4	7	4	4	7	4	4	7
8	16	16	8	16	16	8	16	16	8
9	5	5	9	5	5	9	5	5	9
10	17	17	10	17	17	10	17	17	10
11	6	6	11	6	6	11	6	6	11
12	18	18	12	18	18	12	18	18	12
13	7	7	13	7	7	13	7	7	13
14	19	19	14	19	19	14	19	19	14
15	8	8	15	8	8	15	8	8	15
16	20	20	16	20	20	16	20	20	16
17	9	9	17	9	9	17	9	9	17
18	21	21	18	21	21	18	21	21	18
19	10	10	19	10	10	19	10	10	19
20	22	22	20	22	22	20	22	22	20
21	11	11	21	11	11	21	11	11	21
22	23	23	22	23	23	22	23	23	22
23	12	12	23	12	12	23	12	12	23
24	24	24	24	24	24	24	24	24	24

Fig 3. INTERLEAVING TABLE

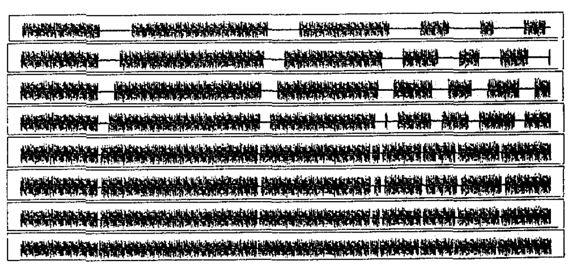

Fig 7. VARYING FADE WIDTH SIGNALS (1 SEC)

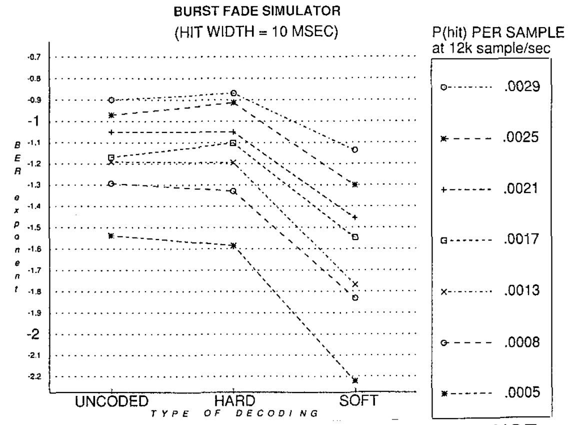

Fig 8. VARYING FADE RATE PERFORMANCE

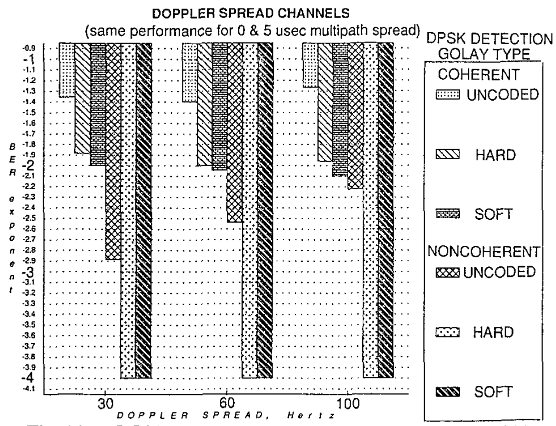

Fig 10. COHERENT VS. NONCOHERENT DPSK

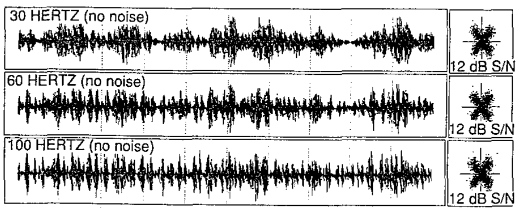

Fig 12. DOPPLER SPREAD SIGNALS (1 SEC)

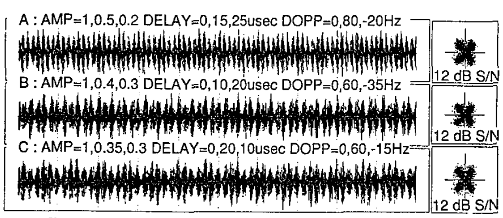

Fig 14. 3-PATH RICIAN SIGNALS (1 SEC)

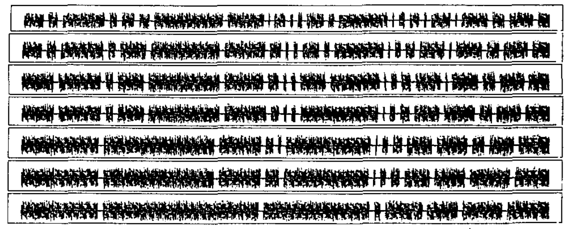

Fig 9. VARYING FADE RATE SIGNALS (1 SEC)

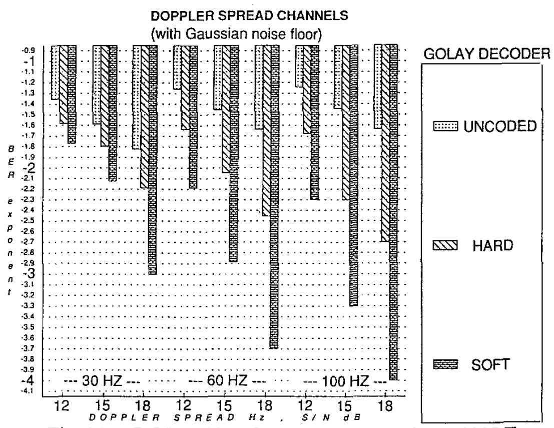

Fig 11. DOPPLER SPREAD PERFORMANCE

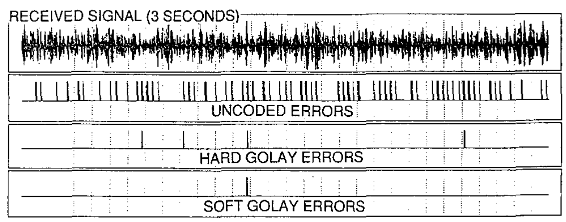

Fig 13. ERROR DISTRIBUTION FOR 60 Hz, 18 dB

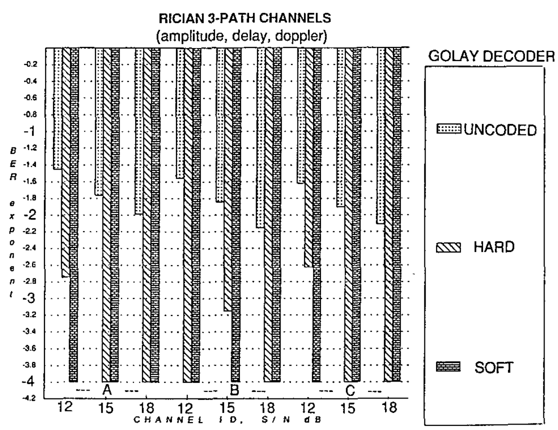

Fig 15. 3-PATH RICIAN PERFORMANCE